



SENSORLESS METHODS FOR DETERMINING THE ROTOR POSITION OF SWITCHED RELUCTANCE MOTORS

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Abstract The attractions of switched reluctance (SR) drives compared to conventional alternatives will be significantly enhanced if the optical rotor position sensor can be eliminated. Considerable attention has recently been applied to various methods for sensorless rotor position measurement, generally based on measurement of phase current and inductance or flux and a pre-knowledge of the magnetic characteristics. This paper comprehensively reviews the various methods previously suggested in both papers and patents for operation both at low and high speeds. Lower speed methods generally utilise the introduction of test signals into the phase winding whilst the winding is normally unenergised. Higher speed methods utilise the main excitation current waveform. In addition there are a number of miscellaneous methods which are considered.

1 INTRODUCTION

The purpose of this paper is to review the main known methods of sensorless position detection for switched reluctance (SR) motors.

To obtain an acceptable performance for an SR motor it is necessary for the excitation of its phases to be carefully timed in relation to rotor position [1]. This has previously necessitated the use of an incremental rotor position transducer which has generally been a disc with teeth or lines together with an optical or electromagnetic sensor which is able to detect the instants the teeth or lines cross prescribed positions.

The incorporation of a rotor position transducer on an SR motor causes additional electrical connections, additional cost and a potential source of unreliability. A considerable variety of methods have therefore been proposed to eliminate the rotor position transducer the majority of which aim to deduce rotor position by the measurement and examination of the current and flux linkage (or inductance) in one or more phases of the motor.

The magnetic characteristics of the motor [2] are represented by the static non-linear relationships between flux linkage $\psi(\theta, i)$ or inductance $L(\theta, i)$ for a motor phase and the phase current i and rotor position θ . Typical characteristics are shown in Figure 1 where θ_1 represents the unaligned rotor position of minimum inductance and θ_n represents the aligned position of maximum inductance. Provided the iron losses in the machine are relatively small the characteristics apply for changing currents and positions - i.e. for dynamic excitation under rotating conditions. The general principle for most sensorless methods is that if at a given instant ψ (or L) and i are measured then θ can be calculated from the pre-stored characteristics. As an alternative, incremental inductance $l(\theta, i)$ can be identified but this is less easy to use - see section 2.1 below.

The methods provide a repetitive measurement of rotor position. These can be effectively continuous with many identifications of position during each phase cycle. Alternatively only one rotor position indicator per phase cycle may be determined generally by finding when the flux or inductance passes through a particular threshold value which represents a particular reference rotor position. This latter case is equivalent to the rotor position sensor signal used at present in most applications. Intermediate positions can be obtained from a phase locked higher frequency signal [3]. In many cases the threshold position is that of minimum or maximum inductance

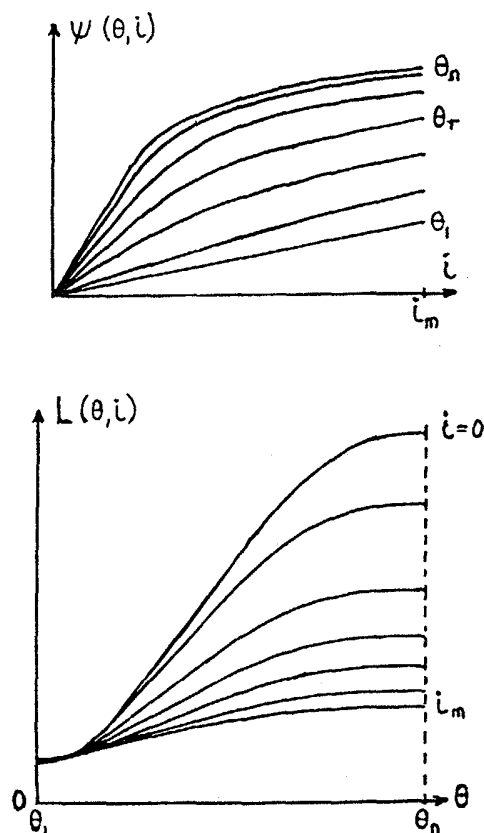


Figure 1 Typical flux linkage $\psi(\theta, i)$ and inductance $L(\theta, i)$ characteristics for an SR motor phase winding.

although the sensitivity with rotor position for these positions is less than during the rising inductance part of the characteristic.

The methods mainly fall into two groups covered below in sections 2 and 3. In the first group test signals of different kinds are introduced during the time when a phase is normally unenergised. For motoring operation this is generally during the falling or, at low speeds, around the minimum inductance periods. The test signals need to be of low amplitude for a number of reasons:

- (a) to avoid negative torque production
- (b) to avoid back-emf effects
- (c) to avoid saturation effects (i.e. dependence of L on i in addition to θ)
- (d) to minimise the size and cost of additional injection circuitry where this is necessary

Reasons (b) and (c) are further discussed in section 2. The low amplitude test signals are susceptible to mutual interference from the excitation currents in other phases, which is the main problem with these methods. Furthermore since at high speeds the excitation waveform occupies the majority of the phase period, injection of test signals is very restricted and hence these methods are more suited for lower speed operation.

The second group of methods utilises the actual excitation current waveform. Typical waveforms for higher speed (the single-pulse mode) and lower speed (the chopping mode) are shown in Figure 2. Sufficiently accurate measurement of total flux-linkage when chopping is difficult (see section 3) and hence these methods are more appropriate for higher speed single-pulse operation. The use of the chopping current waveform is examined in section 2.1.

The sensorless position detection methods reviewed in this paper [4-32] have been limited to rotating SR machines with no additional motor connections for sensing purposes. Some authors (e.g. [33]) have advocated the use of additional phase windings. There have also been many publications on sensorless detection for brushless PM motors which have similarities; some examples are referenced [34-36]. The review also does not consider motor starting in the absence of a discrete position sensor although some authors do mention this. The usual approach [37,38] is to measure the inductance at standstill of at least two phases and from these measurements deduce the active phase for starting in a particular direction.

The aim of the review is to examine and compare the basic principles and ideas of the various methods rather than experimental results. In many cases experimental verification is very limited and no systemic comparison of accuracy has been made. Very often the position detection is an integral part of the control and the verification lies in whether or not the system

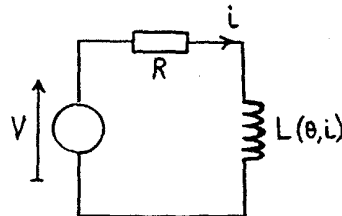


Figure 3 Equivalent circuit for an SR motor phase winding.

R	-	phase resistance
$L(\theta, i)$	-	phase inductance
$\psi(\theta, i)$	-	$L(\theta, i) \cdot i$ - phase flux linkage
i	-	phase current
V	-	phase source voltage

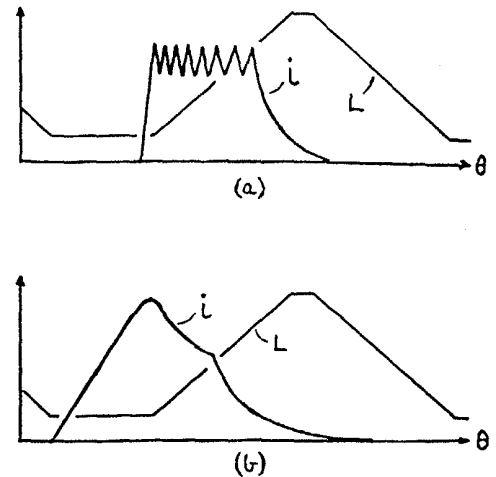


Figure 2 Typical phase currents for:
(a) chopping mode of control
(b) single pulse mode of control
 L represents the idealised inductance profile.

will operate without direct position sensors whilst still producing the same or approximately equivalent performance. Discussion of the actual accuracy of the position detection in terms of degrees of rotor angle has therefore been excluded.

To standardize nomenclature the following symbols are used for test purposes as illustrated by Figure 3.

2 METHODS FOR LOWER SPEED OPERATION

These methods generally make use of an unenergised phase winding for position calculation - i.e. position testing is executed whilst the phase is not required for torque production. Often the tested phase is the one due to be energised next in the sequence and from the tests a position threshold can be established for switching the energisation to this phase. The tests utilise chopping waveforms, or injected high frequency test signals, or a sequence of diagnostic pulses. The current level is relatively small and such that any negative torque produced is negligible and such that the measured inductance $L(\theta, i)$ is effectively the unsaturated inductance $L(\theta, 0)$ and the same as the incremental inductance. Furthermore back emf effects are generally negligible. For example

$$V - Ri = \frac{d\psi(\theta, i)}{dt} = \frac{d(L(\theta, i) \cdot i)}{dt}$$

$$= \left(L(\theta, i) + \frac{\partial L}{\partial \theta}(\theta, i) \right) \frac{di}{dt} + i \frac{\partial L}{\partial \theta}(\theta, i) \frac{d\theta}{dt} = l(\theta, i) \frac{di}{dt} + E_b(\theta, i)$$

where $l(\theta, i) = L + \partial L / \partial i$ is the incremental inductance and $E_b = i(\partial L / \partial \theta) \cdot (d\theta / dt)$ is the back emf. For small i , $\partial L / \partial i$ is negligible.

Since the testing can be effectively continuous provided the frequency is relatively high, continuous position information can be obtained. This may be important at low speeds where excitation current profiling as a function of rotor angle may be required.

However, these methods are generally difficult to implement at high speed since at high speed the excitation current waveform generally occupies the majority of the electrical cycle for a phase and hence there is little unenergised space for position testing.

These methods also suffer from interference due to mutual effects whereby current in the energised phase(s) induce voltages in the testing phase which distort the test current waveform.

2.1 Based on Chopping Waveforms

The first major publication concerning sensorless rotor position detection for SR motors was by Acarnley [4,5] who proposed three methods, the first two of which utilised chopping current waveforms with a fixed current excursion Δi and mean current i . The chopping excursion time is given by

$$\Delta t = \frac{l(\theta, i) \Delta i}{V - Ri - E_b}$$

Provided $V \gg Ri$ and $V \gg E_b$, Δt represents the incremental inductance, which, for constant i is a function of θ . Δt can therefore be compared with a reference Δt_{ref} which represents a reference rotor position. When Δt crosses the reference threshold, energisation is transferred to the next phase in the sequence. Either the current rise time t_r or fall time t_f can be used as Δt , or alternatively the chopping period $t_r + t_f$.

As stated above, two possible methods exist. Firstly the main torque producing chopping waveform may be used. However this suffers from various disadvantages. The range for which position detection is possible is restricted to low speeds else the back emf E_b affects the accuracy. The need for a fixed mean current whilst chopping results in rather inflexible control. Furthermore at high current the incremental inductance can reduce as the rotor and stator align giving a double peaked $l - \theta$ characteristic.

Further work by Panda [6,7] demonstrated that the method was viable but suffered from the restrictions referred to above. To lessen the effect of back emf Panda fixes the commutation point towards the end of the rising inductance period where $dL/d\theta$ is less. This, however, results in appreciable loss of torque as the speed increases, as is shown by his measurements. Panda also attempts with difficulty to extend the method to single pulse control although the outcome appears to be the same as operation synchronously from an oscillator with the rotor taking up an appropriate load angle (see section 4.4).

To avoid the effect of back emf on the relationship between chopping period and rotor angle, Acarnley secondly proposed [4] using a "mini" chopping current waveform in a non-torque productive phase, i.e. when the phase is normally unexcited. This method can be easily implemented in systems utilising chopping or PWM for current profiling in the active phases - the same hardware is used to create a small constant-level current in the inactive phase.

The major difficulty, as previously mentioned, is the mutual coupling with currents in the active phases and this is particularly significant at the higher frequencies associated with PWM. However, a satisfactory elimination of the mutual effect has been found by Egan [8] by using different PWM frequencies for the active torque producing phase and for the probing phase, and by using frequency selective synchronous demodulation of the chopping waveform. The measured inductance profile thereby obtained is free from distortion due to mutual coupling effects. However, a fairly sophisticated demodulation system is required.

2.2 Based on Frequency Injection

As in section 2.1, the idea is to measure inductance in an unenergised phase and to commutate when the inductance exceeds a threshold value. However, rather than using chopping waveforms Ehsani [9] connected the phase winding to an oscillator designed such that the frequency f is inversely proportioned to the phase inductance. Comparison with the threshold value for f is made either using an F to V converter or digitally using a binary counter. The main problem of disconnecting the oscillator from the power circuit during energisation is overcome by using photo-voltaic MOSFET switches. Circuits and details of suitable oscillators are provided in the associated patent [10].

The method requires a separate oscillator for each phase and significant analogue circuitry. It may also be susceptible to corruption of the test signal by mutual effects. The authors recommend measurement around the minimum inductance region where mutual effects are less - however, the sensitivity of inductance to rotor position is also less in this region. Waveforms from a working system are shown for speeds around 1000rpm.

In a later paper [11] Ehsani provides a modification to the frequency injection method. In this case a sinusoidal signal of fixed frequency and amplitude from an oscillator is connected to the phase winding via a resistance. As the inductance of the phase changes, the phase displacement between oscillator voltage and current varies. This can be measured using a phase sensitive demodulator. The current amplitude also varies and may be used as an alternative measurement. It is claimed that this method is extremely robust to switching noise that is present in the sensing phase due to mutually induced voltages.

A further variant to the frequency injection methods is provided by Goetz [12]. The basic circuit for one phase is shown in Figure 4. When the switch S is off a fixed high frequency (f_0) source is connected to a series resonant circuit comprising the phase inductance L and a CR network. The source \tilde{V}_s and coupling capacitor C_1 appear to be common to all phases. The resistor voltage is given by

$$\tilde{V}_R = \frac{\tilde{V}_s}{1 + jQ(f/f_0)^2 - 1}$$

where $f_0 = 1/(2\pi\sqrt{LC})$ and $Q = 1/(2\pi f_0 CR)$. Frequency f_0 is chosen to be slightly greater than the maximum resonant frequency f_{max} , which corresponds to L_{min} . The position at which \tilde{V}_R is a maximum is detected and also corresponds to L_{min} . This method of detecting resonance seems viable in

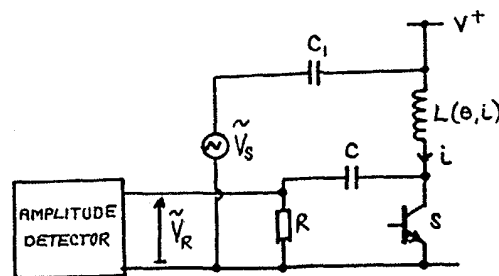


Figure 4 Circuit utilising phase inductance resonance to determine rotor position.

concept but it is unclear how the phase windings are de-energised since there is only one switch per phase and no return current diodes.

The main disadvantage of methods using frequency injection to unenergised phases, apart from limitations at higher speeds and possible mutual effects, is the additional signal processing analogue and digital circuitry required which can be quite significant.

2.3 Based on Diagnostic Pulses

The idea is similar to the previous sections. By using the existing power switching circuit to provide voltage pulses of short duration in an unenergised phase and measuring the consequential current, the inductance $L(\theta, 0)$ can be calculated and θ either determined from L or from the measured flux linkage $\psi(\theta, i)$.

The methods follow from Acarnley's third proposal [4] which is that whilst one phase is being energised, for the next phase in the sequence V is switched on and $\Delta i/\Delta t$ is measured which is a function of θ (since $\Delta i/\Delta t = V/L(\theta, 0)$). If $\Delta i/\Delta t$ is not the correct value then V is turned off and the current allowed to decay to zero. The process is repeated until $\Delta i/\Delta t$ exceeds the threshold value when V is left on and phase changeover occurs.

Acarnley's third method is basically the use of diagnostic pulses in an unenergised phase and has subsequently been investigated and improved by various other researchers. The pulses are generally of a fixed duration Δt at frequencies in the range of 3 to 15 kHz.

Dunlop [13] experimentally investigated the method and found the waveforms for position monitoring were seriously affected by currents in the driving phases and this made absolute rotor position detection very difficult.

McMinn [14,15] proposed measuring the current rise Δi for a fixed Δt and comparing this with a threshold to detect whether the reference position had been reached. There is no difference in principle to measuring $\Delta i/\Delta t$ as proposed by Acarnley. However, McMinn does suggest monitoring variations of the supply voltage V and adjusting the threshold value for Δi accordingly to give the same reference position. He also samples in two phases simultaneously to detect corruption due to switching noise or mutual coupling.

Harris [16] uses the same method but arranges the diagnostic pulses to occur within the 15° period before minimum inductance for a phase - presumably to lessen mutual effects. The relatively large pulses generated significant negative torque. He also discusses the possible use of a state observer in conjunction with this method to improve accuracy (see section 4.1).

Hedland [17] uses an approximate law to compensate for the mutual effect. He uses a diagnostic pulse to measure the apparent inductance, $L_a(\theta, i_a)$ of phase a in the presence of a current i_a in phase b (which is also measured) and corrects this according to the equation

$$L_a(\theta, 0) = (1 + c_{i_b}) L_a(\theta, i_b)$$

$L_a(\theta, 0)$ is then compared with a threshold value to determine the phase commutation point as before.

Mvungi [18, 19] improved the method of correcting for mutual effects to obtain a continuously sampled measurement of position rather than a threshold value. The significant corruption of the test pulses due to mutual coupling is well illustrated by Fig 6 of [19].

The measured flux ψ is corrected for the mutually coupled currents in other phases (i_2, i_3, \dots) according to the linearised law

$$\psi_1(i_1, \theta) = \psi - \left\{ \frac{d\psi}{di_2}(\theta) \cdot i_2 + \frac{d\psi}{di_3}(\theta) \cdot i_3, \dots \right\}$$

where i_1 is the test current and $d\psi/di$ are the mutual coupling coefficients, which, to avoid a multiplicity of look up tables, are taken to be a simplified function of rotor angle. On obtaining the corrected value of flux ψ_1 for current i_1 , θ can be calculated from a 2-dimensional look-up table.

The method was demonstrated to work using a 4-phase 8-pole motor with acceptable position accuracy for speeds up to 1200 rpm on load and 1750 rpm on light load.

3. METHODS FOR HIGHER SPEED OPERATION

Since diagnostic current pulses in an unenergised phase are of necessity small and suffer from mutual effects, it is sensible to examine the use of the main excitation current waveform for the purpose of rotor position sensing. This current already exists and does not require additional switching or injection circuitry.

If at a given instant the flux-linkage $\psi(\theta, i)$ or inductance $L(\theta, i)$ is known and the current i is known, then this defines the rotor position θ provided it is also known whether the inductance is rising or falling. The latter is generally obvious from the positioning of the excitation which will be predominantly in the rising inductance region for motoring operation. The position can be looked up in pre-stored tables of ψ or L against θ and i .

The main problem is accurately measuring the flux by integration of $V-Ri$. For single pulse operation this is not difficult and if $V \gg Ri$ integration of voltage or even multiplication of V by time may suffice to obtain sufficient accuracy. However, under chopping conditions with repetitive reversals of the phase voltage over a relatively longer time the Ri effect is very significant since it provides volt-seconds continually of the same polarity whereas the positive and negative excursions of V are largely self-cancelling. For this reason these methods are considered more suited to higher speed single pulse operation.

The first proponent of this method appears to be Hedland [20] who was able to simplify the idea to avoid the use of multi-dimensional look up tables. The aim is to identify a particular rotor position θ_{ref} for each phase where θ_{ref} corresponds to a point in the rising inductance region (see Fig. 1). Values of ψ against i are stored for the particular θ_{ref} . From the commencement of energisation for a phase, the current i and flux ψ are continuously sampled and the flux $\psi_{ref}(\theta_{ref}, i)$ is calculated - i.e. the flux for i if the position θ was equal to θ_{ref} . Initially the measured flux $\psi < \psi_{ref}$ but eventually, when θ passes θ_{ref} , ψ becomes $> \psi_{ref}$. This transition is detected to identify the reference or threshold position. A rotor position indicator is thus obtained for each phase period from which the excitation can be timed. For acceptable resolution in defining θ_{ref} at high speeds a high sampling period is required. The sampling and comparison may therefore require a dedicated microprocessor or equivalent circuitry.

Exactly the same method is proposed by Lyons [21 - 25] with the measurement regime restricted to a 60° electrical band (for a 3-phase motor) at the beginning of the rising inductance period or the end of falling inductance [21, 24]. Some additional features are proposed. Firstly [21,22] mutual effects can be corrected by using a multi-dimensional look up table or map for calculating $\psi_{ref} = f(I_1, I_2, \dots)$ using constant mutual

coupling coefficients where I_1 is the current on the active phase and I_2, \dots are the other phase currents. Alternatively [21, 25] a lumped parameter reluctance network model of the motor including all the mutual coupling effects can be used. The network is very complex and many of the reluctance elements are functions of rotor angle θ so this, in the authors opinion, would be difficult to implement in practice. A method [23] of checking whether the measurement is within bounds - i.e. the position detection sequence is in lock - is also proposed. The position estimation algorithm was tested on an off-line basis (i.e. from recorded measurements of voltage, current, speed and rotor position) and is claimed to give good agreement between measured and estimated angles.

4 OTHER METHODS

The previous sections cover the main-stream methods for sensorless detection of SR rotor position. However, there are a number of other proposals and published work which do not directly fit into these sections.

4.1 Use of a State Observer

This has been proposed by Lang [26] and is illustrated in Figure 5. A mathematical model of the complete system including the mechanical load is provided in state-space format. The inputs to both model and actual motor are the phase voltages \underline{v} . The model assumes a load of constant inertia J and viscous friction B and an inverse inductance function $H(\theta) = (L(\theta))^{-1}$ which is independent of current such that torque $= \frac{1}{2} \psi^2 dH/d\theta$ for a given phase. The model enables values of the currents \underline{i} , flux linkages $\underline{\psi}$, speed w and position θ to be estimated. The estimated currents are compared with the measured motor currents and the current errors are used to adjust the values of J, B and $H(\theta)$ using an adjustment matrix. If the predicted currents match the real currents then it is implied that the model rotor position θ will be correct. This is not necessarily true although by matching the currents the error in predicting θ would be reduced.

The success of the method must largely depend on the accuracy of the model. In many applications the load will be non-linear in addition to the motor and a more sophisticated model is probably necessary.

The amount of computation on a repetitive basis is also large, and with a more sophisticated model it is questionable whether the computation would be sufficiently fast.

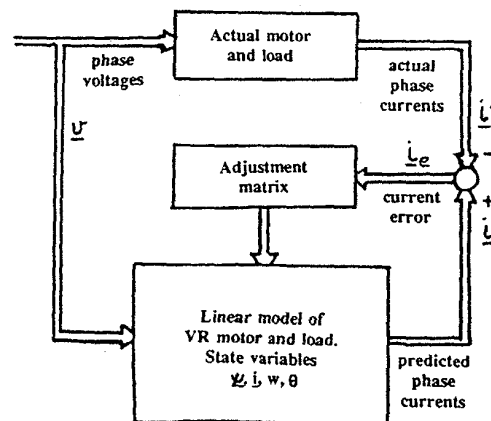


Figure 5 System using state observer for rotor position estimation.

4.2 Monitoring Back emf

An early idea by Piatkowski [27] is to supply the phase winding with a constant bias current whilst it is normally unenergised and to monitor the resultant back emf $L di/dt$. At the middle of the minimum inductance region the back emf will change polarity and this can be used as a position indicator. The application was a small single phase 12V pump motor. For multiphase motors the mutual emf from other phases would be a problem. Furthermore early turn-on in the single phase control mode (in advance of minimum inductance) would not be possible.

4.3 Monitoring Mutually Induced Voltages

A recent idea by Austermann [28] is to monitor the mutually induced voltage u_1 in an unenergised phase winding due to current i_2 in an energised phase.

$$u_1 = \frac{d}{dt} \psi_m(i_2, \theta) = \frac{\partial \psi_m}{\partial i_2} \frac{di_2}{dt} + \frac{\partial \psi_m}{\partial \theta} \frac{d\theta}{dt}$$

where m is the mutually induced flux linkage. In the chopping mode with substantially constant mean current the $(\partial \psi_m / \partial i_2) di_2 / dt$ term is at relatively high frequency and can be filtered out, whereas $\partial \psi_m / \partial \theta$ is at phase frequency ($d\theta/dt$ is effectively constant over the measurement time) and is claimed to pass through zero at an accurately prescribed position presumably independent of i_2 . A position indicator is thereby obtained.

Ehsani [29] has used substantially the same method as Austermann and claims good results over a wide speed range. The method appears to be limited to the chopping mode.

4.4 Synchronous Control Methods

A further group of methods for sensorless SR motor control [30-32] do not provide a direct position indicator. The motor is predominantly run from a variable frequency oscillator in the conventional synchronous manner, but adjustment is made to the dwell angle or frequency to give improved stability. The resulting control is less flexible and the performance poorer than with systems for which the control is directly based on rotor position measurement. Publications predominantly concern control rather than position measurement but in so far that these methods enable operation of an SR motor without a rotor position sensor they are alternative contenders to the other methods presented in this review.

Miller [30] adjusts the dwell angle (or ON-angle) for variation in load torque whilst keeping the torque angle (the angle at which commutation occurs) fixed. The dwell angle is adjusted by a feedback loop dependent on measurement of mean supply current. The resulting control appears to be very underdamped but the stability may be improved if the speed is allowed to droop with load. The method is apparently aimed at operation at a single rather than variable speed.

Vukosavic's method [31] is similar except that the dwell angle is fixed and the advance angle is varied rather than vice-versa. To achieve this aim a feedback signal, based on measurement of the mean energy-return current (i.e. the current in the power converter diodes), is used to adjust the oscillator frequency. Some speed droop is therefore inherent in the performance. (8% at full load)

Obravodic's method [32] differs in that the energisation is by chopping rather than single pulse control - the torque is adjusted by chopping current level rather than by dwell angle or advance angle. As in Acarnley's third method, the initial Δi

for a fixed Δt is measured and compared with a reference Δi_{ref} . The error signal $\Delta i_{ref} - \Delta i$ is used to raise or lower the chopping current level. Feedback loop compensation may be necessary to give acceptable stability.

The relevance of these methods is that it is not absolutely necessary to have a position reference signal for SR motor control. The disadvantage is that the stability and performance are inevitably poorer, but there is the advantage that the controller will be relatively less expensive.

5. CONCLUSION

It is evident from the number of publications that research activity in the area of sensorless position detection is widespread and considerable further work will be required before a reliable and commercially applicable method is fully developed. It is also evident that either significant computation is required to obtain the information from existing or readily created waveforms, or significant additional circuitry is required to implement the less sophisticated schemes. It is possible that the most appropriate method may differ according to the application - for example, a more expensive drive with high dynamic performance as compared with a simpler drive for domestic or pump/fan applications. Of the methods reviewed in this paper, none appeared to be suitable for a wide range of speeds over which the power capability of the motor is fully utilised, and it is possible that a combination of methods may be necessary in this case.

The effect of iron losses on some of the methods may need investigation - some authors [e.g. 8] have mentioned these, but detailed study has been outside the scope of most work so far.

Much more extensive testing and published results of performance and positional accuracy will be necessary before a preferred method (if any) will be apparent. It is also likely that improvements will be made to the existing methods and that further new methods will be forthcoming.

In conclusion therefore, the state of this art is still evolving with significant interest from a large number of researchers and the authors of this paper would hesitate to recommend any particular method at this stage.

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Note: For patents, the year specified is that of first filing or the priority date.

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